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Interference Cancellation
in
Wireless Communications

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To my parents, Dimitra and Konstantinos.

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Abstract

One of the most significant problems in wireless communications is interference. Indeed, there are a lot of different kinds of interference. In this thesis, we consider point-to-point systems which are interfered from unwanted sources which communicate simultaneously over the same frequency band.

At the beginning, we study the "Reciprocity" Property and we check whether the channel reciprocity property holds true in practice. Using the Universal Software Radio Peripherals (USRPs), we observe that it does not hold true, due to the non-symmetric characteristics of the RF electronic circuits.

Then, we study a scenario which includes two co-existing and interfering point-to-point wireless links. Each transmitter, knowing the channels to its unintended receiver, is able to pre-cancel its induced interference. We verify the related concepts by implementing the channel feedback and beamforming techniques on a USRP testbed, and we observe interference cancellation at the unintended receivers.

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Chapter 1

Introduction

Wireless communication is the transfer of information between two or more points without using wires, using e.g. electromagnetic waves. The information is carried through the channel that is formed between the communication antennas.

1.1 The interference problem

One of the most significant problems in wireless communications is *interference*. There are different types of interference: Electromagnetic interference (EMI), Co-channel interference (CCI), Adjacent-channel interference (ACI), Intersymbol interference (ISI), Common-mode interference (CMI), Conducted interference, etc. In our study, we consider the interference induced from systems that communicate simultaneously over the same frequency band.

1.2 Thesis Outline

Our target is to study a scenario with two point-to-point wireless links, that communicate simultaneously over the same frequency band and suggest techniques for interference cancellation.

At first, we study and implement, on a USRP testbed, a single-antenna point-to-point system. Then, we implement a two-input/one-output and an one-input/two-output system. By studying the "uplink" and "downlink" channels, we investigate whether the channel reciprocity property holds true. We conclude that it does not hold true in our case. Finally, in order to develop techniques for interference cancellation we adopt a channel feedback approach. Thus, the transmitters knowing their channels to their unintended receivers are able to pre-cancel the interference caused to the unintended receivers by appropriate beamforming.

We verify our theoretical approach by implementing the channel feedback and beamforming techniques on a USRP testbed.

Chapter 2

Software Tools and USRPs

2.1 Software Defined Radio (SDRs)

A software-defined radio system, or SDR, is a radio communication system where components that are typically implemented in hardware (e.g. mixers, filters, amplifiers, modulators/demodulators, detectors, etc.) are instead implemented in software on a personal computer or embedded system.

A basic SDR system may consist of a personal computer equipped with a sound card, or other analog-to-digital converter, preceded by some form of RF front end. Significant amounts of signal processing are handed over to the general-purpose processor, rather than being done in special-purpose hardware. Such a design produces a radio which can operate under different protocols.

2.2 Universal Software Radio Peripheral (USRP)

The Universal Software Radio Peripheral (USRP) [3] products are computer-hosted software radios. They are designed and sold by Ettus Research, LLC and its parent company, National Instruments. The USRP product family is intended to be a comparatively inexpensive hardware platform for software radio, and is commonly used by research labs, and universities. USRPs connect to a host computer through a high-speed USB or Gigabit Ethernet link. Some USRP models integrate the general functionality of a host computer with an embedded processor that allows the USRP Embedded Series to operate in a standalone fashion.

2.2.1 USRP 1

The USRP1 is the original Universal Software Radio Peripheral hardware (USRP) that provides entry-level RF processing capability. It is intended to provide software defined radio development capability for cost-sensitive users and applications. The architecture includes an Altera Cyclone FPGA, on 64 MS/s dual ADC, on 128 MS/s dual DAC and USB 2.0 connectivity to provide data to host computers. A modular design allows the USRP1 to operate from DC to 6 GHz. The USRP1 platform can support two complete RF daughterboards. This feature makes the USRP ideal for applications requiring high isolation between transmit and receive chains, or dual-band transmit/receive operation. The USRP1 can stream up to 8 MS/s to and from host applications, and users can implement custom functions in the FPGA fabric.

RFX2400 daugterboard

This daughterboard works in the 2.4Ghz band. This band consists of a continuous spectrum range of 100 Mhz (one of the areas of the ISM band). The daughterboard has one transmitter and one receiver. The transmitter takes the baseband analog signal which comes from the mainboard and modulates it to the central frequency that we choose through the software (gnu radio). Pulse shaping is also implemented in the software. The output of the transmitter goes to a two-sided switch which is connected to the input of the transmitter from the one side and to the antenna plug from the other side. The side of the switch is controlled through the software, depending on whether we transmit or receive. The receiver consists of an oscillator whose frequency is controlled by the software, and a mixer which mixes the signal coming from the antenna with the sinusodial signal coming from the oscillator. The receiver is direct conversion so the baseband signal which is produced is sent to the mainboard for sampling.

Mainboard

The USRP has 4 high-speed analog to digital converters (ADCs), each at 12 bits per sample, 64MSamples/sec. There are also 4 high-speed digital to analog converters (DACs), each at 14 bits per sample, 128MSamples/sec. These 4 input and 4 output channels are connected to an Altera Cyclone EP1C12 FPGA. The FPGA, in turn, is connected to a USB2 interface chip, the Cypress FX2, and to the computer. The USRP is connected to the computer via a high speed USB2 interface. So, in principle, we have 4 input and 4 output channels if we use real sampling. However, we can have more flexibility (and bandwidth) if we use complex (IQ) sampling. Then we have to pair them up, so we get 2 complex inputs and 2 complex outputs.

ADC

There are 4 high-speed 12-bit AD converters. The sampling rate is 64M samples per second. In principle, each ADC can digitize a band as wide as 32MHz. The full range of the ADCs is 2V peak to peak, and the input is 50 ohms differential. This is 10mW, or 10dBm. There is a programmable gain amplifier (PGA) before the ADCs which amplifies the input signal so that it utilizes the entire input range of the ADCs, in case the signal is weak. The PGA is up to 20dB. With gain set to zero, full scale inputs are 2 Volts peak-to-peak differential. When set to 20 dB, only .2 V p-p differential input signal is needed to reach full scale. This PGA is software programmable.

DAC

At the transmit path, there are also 4 high-speed 14-bit DA converters. The DAC clock frequency is 128 MS/s, so Nyquist frequency is 64MHz. However, we will probably want to stay below it to make filtering easier. A useful output frequency range is from DC to about 44MHz. The DACs can supply 1V peak to a 50 ohm differential load, or 10mW (10dBm). There is also PGA used after the DAC, providing up to 20dB gain. This PGA is software programmable. The DAC signals $IOUTP_A/IOUTN_A$ and $IOUTP_B/IOUTN_B$ are current-output, each varying between 0 and 20 mA. They can be converted into differential voltages with a resistor.

FPGA

According to the above, the information rate that a USRP sends and receives at the same time is:

$$(64MSPS * 12bit/Sample + 128MSPS * 14bit/Sample) * 2 = 640Mbyte/sec$$

But the data rate that the USB port can support is up to 32Mbyte/sec. Moreover the samples should be transformed from 12 and 14 bits to the closest multiple of 8 bits. These processes are undertaken by the FPGA. On the FPGA, there are two digital filters (decimation and interpolation). The decimation filter's input is the flow of the samples from the ADC which comes with a rate of 64MS/s. The decimation filter subsamples its input flow by a decimation rate factor, which is chosen through the software. Accordingly, the interpolation filter oversamples the data flow by an interpolation rate factor, which is also chosen through the software, so that the final sampling rate is 128MS/s (sampling rate of the DAC). Finally, as the data rate of the USB is not stable, we need a buffer before the interpolation filter in which the samples coming from the PC are stored and they are read by a rate of 128MSPS/interpolation so that in the end we have a 128MS/s.

2.3 GNU radio

GNU Radio is a free and open source software development toolkit that provides signal processing blocks to implement software-defined radio systems. It can be used with readily-available low-cost external RF hardware to create software-defined radios, or without hardware in a simulation environment. It is widely used in academic and commercial environments to support both wireless communications research and real-world radio systems. GNU Radio applications are primarily written using the Python programming language, while the supplied performance-critical signal processing path is implemented in C++ using processor floating-point extensions, where available. Thus, the developer is able to implement real-time, high-throughput radio systems in a simple-to-use, application-development environment. GNU Radio supports development of signal processing algorithms using pre-recorded or generated data, avoiding the need for actual RF hardware.

Chapter 3

Single Antenna Point to Point System

In this section, we consider a packet-based single-antenna Point-to-Point system.

3.1 Transmitter

Each transmit packet consists of N complex 4-QAM (information and training) symbols. Trainning symbols are located at specific positions, as depicted in Figure 3.1. The transmitter creates the communication packet which is oversampled,

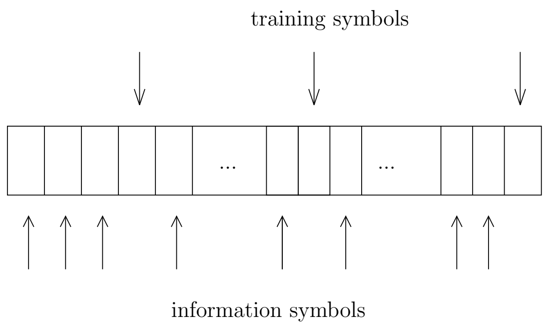

Figure 3.1: Packet with information and training symbols

passed through an oversampled Square Root Raised Cosine (SRRC) transmit filter, amplified and modulated. Thus, the transmitter sends:

$$u(t) = \Re\left\{\sum_{i=0}^{N-1} a_i g_T(t - iT)e^{-j2\pi F_c t}\right\},$$

where a_i are the transmit symbols, T is the symbol period, F_c is the modulation frequency and g_T the SRRC filter.

3.2 Receiver

The first task of the receiver is to detect the packet and to get synchronized in time. In our implementation, we have used the energy of the received signal as the metric for packet detection and symbol synchronization.

Carrier Frequency Offset (CFO)

The receiver demodulates the packet using the local carriers $\cos(2\pi(F_c+\Delta F)t+\phi)$ and $-\sin(2\pi(F_c+\Delta F)t+\phi)$. Thus, during the demodulation process frequency carriers ΔF and $2F_c+\Delta F$ are generated. The carrier ($2F_c+\Delta F$) is suppressed by the lowpass filter of the receiver. However, the carrier ΔF can cause significant problems. Thus, it must be estimated and corrected. The received signal has the form:

$$r(t) = he^{-j(2\pi\Delta F t+\phi)}\sum_{i=0}^{N-1} a_i g_T(t - iT) + n(t),$$

where h is the flat fading channel and $n(t)$ is Additive White Gaussian Noise (AWGN). After matched filtering and sampling, the symbol-spaced received sequence becomes:

$$r_k = ha_k e^{-j(2\pi\Delta F(kT)+\phi)} + n_k, \quad k = 0, 1, .., N-1.$$

Since the receiver knows the training symbols a_{lm} (m positive integer and $l = 0, 1, ..., N_{tr}-1$), we can estimate the above Carrier Frequency Offset (CFO) using the trainning sequence. Our estimate is the frequency f_* at which the Fourier

Tranform of the sequence $c_l = hr_{lm}a^*_{lm} = h|a_{lm}|^2 e^{-j(2\pi\Delta FTlm+\phi)}$ is maximized. Finally, we cancel the CFO, by forming the sequence:

$$z_k = e^{j(2\pi \frac{f_*}{m} k)} r_k = ha_k + n_k,$$

where we have assumed perfect CFO estimation.

Channel Estimation and correction

The flat fading channel multiplies the signal by an unknown complex number $h = h_r + jh_i$. Using the training symbols we estimate the channel as follows:

$$\hat{h} = \frac{1}{N_{tr}} \sum_{l=0}^{N_{tr}-1} \frac{z_{lm}}{a_{lm}}.$$

In order to perform coherent detection, we multiply the sequence z_k by $\frac{\hat{h}^*}{|\hat{h}|^2}$. Assuming perfect channel estimation, that is $\hat{h} = h$, we obtain:

$$y_k = \frac{h^*}{|h|^2} z_k = \frac{h^*}{|h|^2}(ha_k + n_k) = a_k + n'_k.$$

Channels

In order to see how the channels look like in a static environment, we estimated the channels for 1000 consecutive packets. The idle period between consecutive packets is set to 100 msec. In Figure 3.2, we plot the channel estimates. We observe that the channel estimates are uniformly distributed on a circle of the complex plane. This happens because their magnitude does not change, due to the static environment. However, each channel has a random phase, which seems to be uniformly distributed in $[0, 2\pi)$.

3.3 "Ping-Pong"

In this section, we discuss the implentation issues of a "Ping-Pong" protocol. The basic aspects of this protocol are as follows:

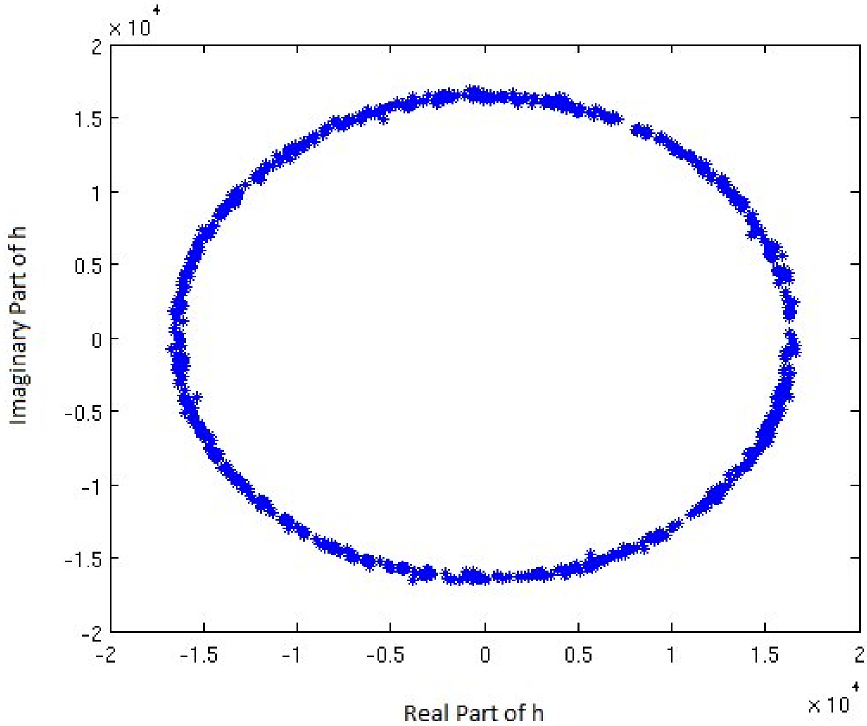

Figure 3.2: Channel Estimations

- There are two distinct nodes (USRPs); we denote them as node A and node B. Each node alternates between the Receiver (RX) and the Transmitter (TX) mode.
- When a node is at TX mode, it sends N packets and then it changes to RX mode.
- When a node is at RX mode, it waits in order to receive $N_1 \leq N$ packets, then waits a few msec and finally changes to TX mode.
- We call the channel from node A to node B as *uplink* channel, and the channel from node B to node A as *downlink* channel.

Chapter 4

Multiple Antenna Point to Point Systems

In this chapter, we consider point-to-point systems where one of the two nodes has two antennas. More precisely, we consider a 2x1 system which has a two-antenna transmitter and a single-antenna receiver, and then consider a 1x2 system, which has a single-antenna transmitter and a two-antenna receiver.

4.1 A 2x1 system

The transmitter has two antennas and sends two sequences; one sequence per antenna, as depicted in Figure 4.1. We assume that some symbols are known to the receiver i.e. these symbols are training symbols.

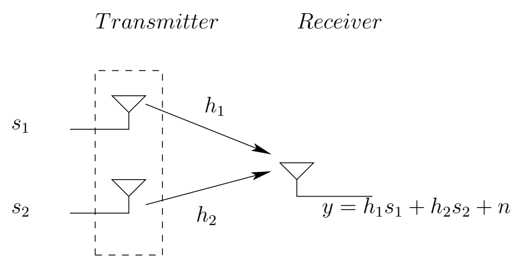

Figure 4.1: 2x1 System

After matched filtering, frame and symbol synchronization, and considering only the training symbols, the received sequence can be expressed as:

$$\begin{bmatrix} r_1 \\ r_2 \\ \vdots \\ r_{N_{tr}} \end{bmatrix} = \Gamma(u) \begin{bmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \\ \vdots & \vdots \\ s_{N_{tr}1} & s_{N_{tr}2} \end{bmatrix} \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \\ \vdots \\ n_{N_{tr}} \end{bmatrix}$$

or equivalently,

$$\boxed{r = \Gamma(u)Sh + n,}$$

where N_{tr} is the length of the training sequence, Γ is the matrix

$$\Gamma(u) = \text{diag}\{1, e^{j2\pi u}, e^{j4\pi u}, ..., e^{j2\pi(N_{tr}-1)u}\},$$

$u = \Delta FT$ and T is the symbol period.

Our aim is to estimate and cancel u. An estimate of u is provided by the maximizing argument of the function [4]:

$$g(\tilde{u}) = r^H \Gamma(\tilde{u}) B \Gamma^H(\tilde{u}) r$$

written as,

$$\hat{u} := \arg\max_{\tilde{u}} g(\tilde{u}),$$

where $B = S(S^H S)^{-1} S^H$. Then, we can cancel the influence of the CFO if we multiply the received signal r_k with the exponential signal $e^{-j2\pi\hat{u}k}$, for $k = 1, 2, ..., N$.

Finally, we can estimate the *channel matrix* h, as [4]:

$$\hat{h}(\hat{u}) = (S^H S)^{-1} S^H \Gamma^H(\hat{u}) r.$$

4.2 A 1x2 system

This system has a single-antenna transmitter and a two-antenna receiver, as depicted in Figure 4.2. The transmitter sends a sequence s. We assume that some symbols are known to the receiver. Assuming perfect synchronization and

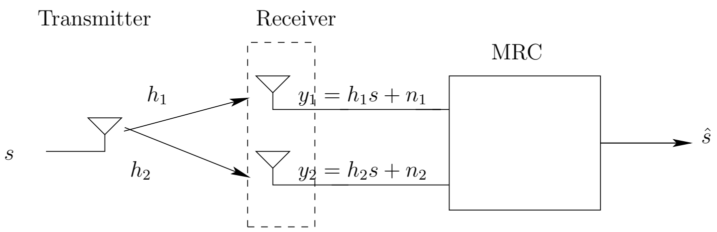

Figure 4.2: 1x2 System

CFO estimation-cancellation, the sequence received at the two-antenna receiver can be expressed as:

$$\begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} s + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix},$$

or, in vector notation:

$$\underline{y} = \underline{h}s + \underline{n}.$$

We can estimate each of the channels h_1 and h_2 using the same technique as in the single antenna point-to-point system. Finally, assuming perfect channel knowledge, we can apply the Maximum Ratio Combiner (MRC) and get:

$$\underline{r} = \frac{\underline{h}^*}{||\underline{h}||^2}\underline{y} = \frac{\underline{h}^*}{||\underline{h}||^2}(\underline{h}s + \underline{n}) = s + \underline{n}'.$$

Then, we can proceed to demodulation.

4.3 "Ping-Pong" - Channel Feedback

The "Ping-Pong" protocol has the same characteristics as before. The new issues are:

- There are two *uplink* channels, and two *downlink* channels. These channels are presented at Figure 4.3.

- After k received packets, the receiver estimates the channels h_1 and h_2, and then sends its estimates (channel feedback) to the other node.

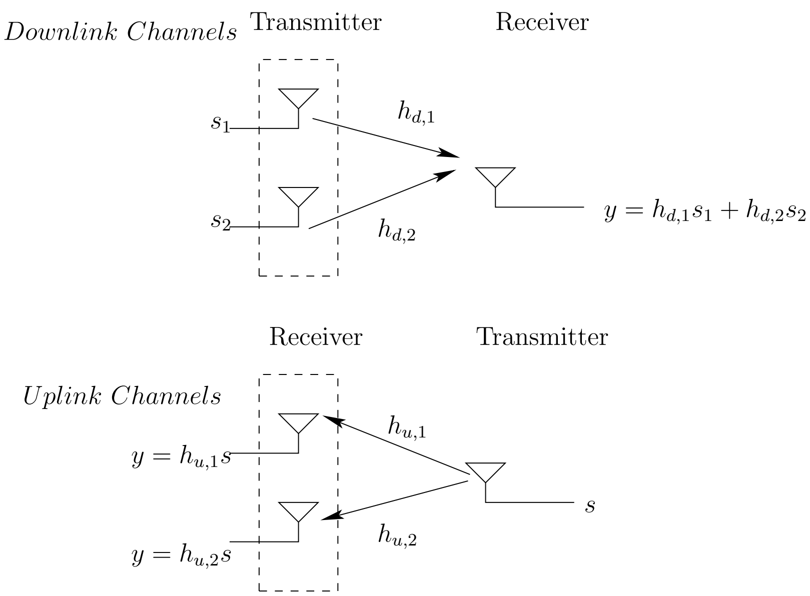

Figure 4.3: Uplink and Downlink Channels

Chapter 5

Channel Reciprocity

5.1 Introduction

In this chapter, we study the channel Reciprocity Property. We assume that both the forward and the reserve links occur at the same frequency. Let us denote the two different transceivers as A and B.
Since wireless communication systems are often full-duplex, the reciprocity principle suggests that the transceiver A can obtain the forward (A to B) channel from the reserve (B to A) channel measumerements, as depicted in Figure 5.1.

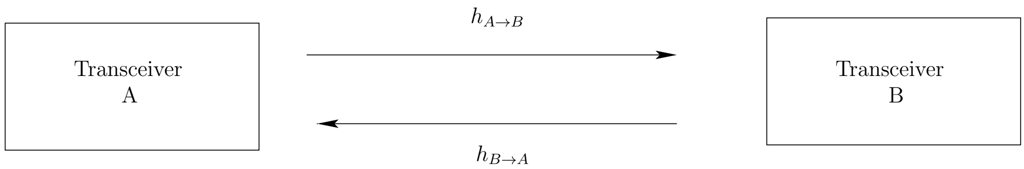

Figure 5.1: Reciprocity Property

The reciprocity property is based on the fact that electromagnetic waves traveling in both directions undergo the same physical distructions *(i.e. reflection, refraction, diffraction, etc)*. Therefore, if the links operate at the same frequency band in both directions, the impulse response of the channel observed between any two antennas should be "the same" regardless of the direction. Despite the fact that the electromagnetic foundations of the reciprocity principle, due to H.

A. Lorentz, have been known since 1896 and extensively explored, applications in the field of wireless communications have been scarce. This is due to the general understanding that the non-symmetric characteristics of the radio-frequency (RF) electronic circuitry may break the reciprocity property.

We can check whether the wireless channels are reciprocal or not, based on the channel estimates.

Channels

In this section, we present the two uplink and two downlink channel estimates, for 1000 packets (assuming that the environment is *static*). We plot the four channels in Figure 5.2. Indeed, the four channels form four distinct concentric

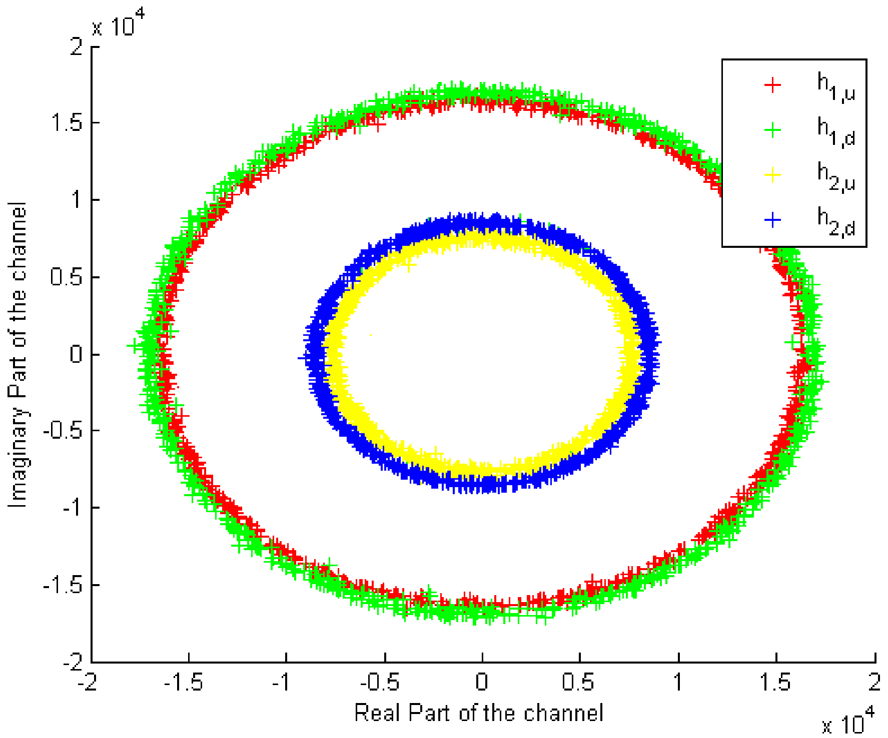

Figure 5.2: Channel Estimations

circles. We can clearly observe two different circles for the channel h_1, and two different circles for the channel h_2; the two different circles for the same channel represent the uplink and the downlink estimates for a particular channel. Thus, the uplink and downlink estimates for the same channel (either the channel h_1 or the channel h_2) are close enough as it regards their radius (*norm of the channel*).

5.2 Reciprocity

Let us assume a two-antenna node, termed the Base Station (**BS**), and a single-antenna node, termed the User.
We denote the channel from the Base Station to the User, as the Downlink Channel $h_d = \begin{bmatrix} h_{d,1} \\ h_{d,2} \end{bmatrix}$, and the channel from the User to the Base Station, as the Uplink Channel $h_u = \begin{bmatrix} h_{u,1} \\ h_{u,2} \end{bmatrix}$, as depicted in Figure 5.3. Moreover, we

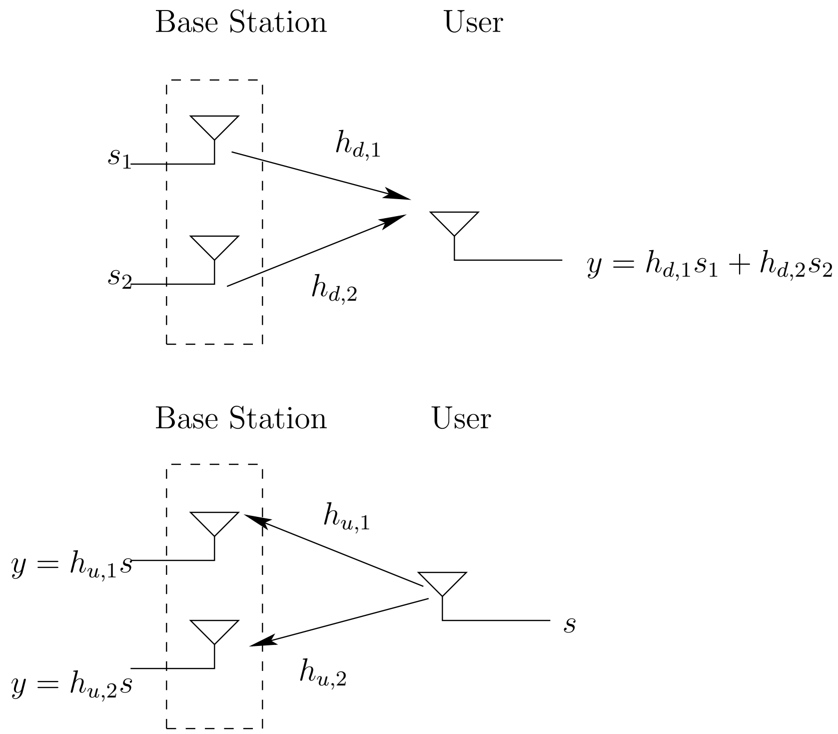

Figure 5.3: Base Station and User

consider the following expression:

$$\boxed{\begin{bmatrix} h_{u,1} \\ h_{u,2} \end{bmatrix} = e^{j\phi_{u,d}} \begin{bmatrix} h_{d,1} \\ h_{d,2} \end{bmatrix}}$$

where $e^{j\phi_{u,d}}$ is the assumed common phase difference of the two channels. If the above expression holds true then the channels are termed reciprocal.

At first, we check whether there is a *common* phase difference between two Uplink or Downlink channels at two different time slots or not. Finally, we check if there is a common phase difference between an Uplink and a Downlink channel.

5.2.1 Uplink Paths

We can check if there is a common phase difference between two uplink channel estimates at two different times slots t_1 and t_2, by checking if the following expression holds true:

$$\begin{bmatrix} h_{u_{t_1},1} \\ h_{u_{t_1},2} \end{bmatrix} = e^{j\phi} \begin{bmatrix} h_{u_{t_2},1} \\ h_{u_{t_2},2} \end{bmatrix} \tag{5.1}$$

where

- $h_{u_{t_1},i}$ are the uplink channels $i = 1, 2$, at the time slot t_1,
- $h_{u_{t_2},i}$ are the uplink channels $i = 1, 2$, at the time slot t_2,
- $e^{j\phi}$ is the assumed common phase difference for the uplink channels, at time slots t_1 and t_2.

Let $r_{t_1,t_2,i}$ be the ratio:

$$r_{t_1,t_2,i} = \frac{h_{u_{t_1},i}}{h_{u_{t_2},i}}, \quad i = 1, 2.$$

Considering the time slots t_i and t_j, we define the ratio $R_{i,j}$ as follows:

$$R_{i,j} = \frac{r_{t_i,t_j,1}}{r_{t_i,t_j,2}} = \frac{\dfrac{h_{u_{t_i},1}}{h_{u_{t_j},1}}}{\dfrac{h_{u_{t_i},2}}{h_{u_{t_j},2}}}$$

If expression (5.1) holds true, then $R = 1$, otherwise $R \neq 1$.

Experiments

In this section, we use the uplink channel estimates at different time slots, and check if there is a common phase different between them. We follow the procedure described in the previous section, and calculate the ratios $R_{i,j}$ for various t_i, t_j. In Figure 5.4, we plot the ratios $R_{i,j}$ for $i = 1, 2, ..., 99$ and $j = 2, 3, .., 100$ (or, more precisely $R_{1,2}, R_{2,3}, .., R_{99,100}$).

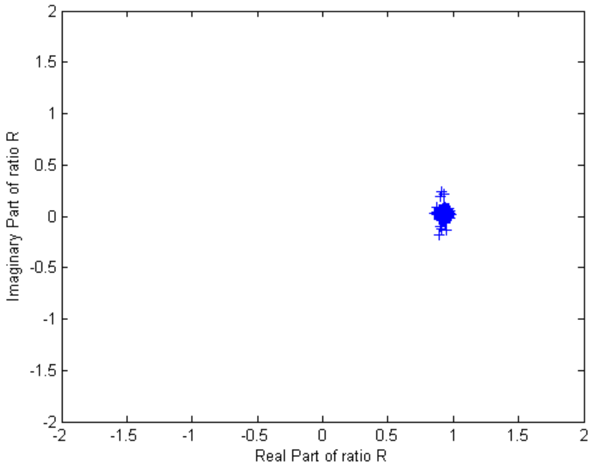

Figure 5.4: Ratios R for different uplink channel estimates

We observe that the ratios $R_{i,j}$ are located close to the point (1,0). This means that the 2x1 vector channels have a common phase difference, i.e. expression (5.1) holds true.

5.2.2 Downlink Paths

We can check if there is a common phase difference between two downlink channel estimates at two different times slots t_1 and t_2, by checking if the following expression holds true:

$$\begin{bmatrix} h_{d_{t_1},1} \\ h_{d_{t_1},2} \end{bmatrix} = e^{j\phi} \begin{bmatrix} h_{d_{t_2},1} \\ h_{d_{t_2},2} \end{bmatrix} \tag{5.2}$$

where

- $h_{d_{t_1},i}$ are the downlink channels $i = 1, 2$, at the time slot t_1,

- $h_{d_{t_2},i}$ are the downlink channels $i = 1, 2$, at the time slot t_2,
- $e^{j\phi}$ is the assumed common phase difference for the downlink channels, at time slots t_1 and t_2.

Let $r_{t_1,t_2,i}$ be the ratio:

$$r_{t_1,t_2,i} = \frac{h_{d_{t_1},i}}{h_{d_{t_2},i}}, \quad i = 1, 2.$$

Considering the time slots t_i and t_j, we define the ratio $R_{i,j}$ as follows:

$$R_{i,j} = \frac{r_{t_i,t_j,1}}{r_{t_i,t_j,2}} = \frac{\dfrac{h_{d_{t_i},1}}{h_{d_{t_j},1}}}{\dfrac{h_{d_{t_i},2}}{h_{d_{t_j},2}}}$$

If expression (5.2) holds true, then $R = 1$, otherwise $R \neq 1$.

Experiments

In this section, we use the downlink channel estimates at different time slots, and check if there is a common phase different between them. We follow the procedure described in the previous section, and calculate the ratios $R_{i,j}$ for various t_i, t_j. In Figure 5.5, we plot the ratios $R_{i,j}$ for $i = 1, 2, ..., 99$ and $j = 2, 3, .., 100$ (or, more precisely $R_{1,2}, R_{2,3}, .., R_{99,100}$).

We observe that the ratios $R_{i,j}$ are located close to the point (1,0). This means that the 2x1 vector channels have a common phase difference, i.e. expression (5.2) holds true.

5.2.3 Up & Downlink Reciprocity

We can check if there is a common phase difference between an uplink and a downlink channels estimation, by checking if the following expression holds true:

$$\begin{bmatrix} h_{u,1} \\ h_{u,2} \end{bmatrix} = e^{j\phi} \begin{bmatrix} h_{d,1} \\ h_{d,2} \end{bmatrix} \tag{5.3}$$

where,

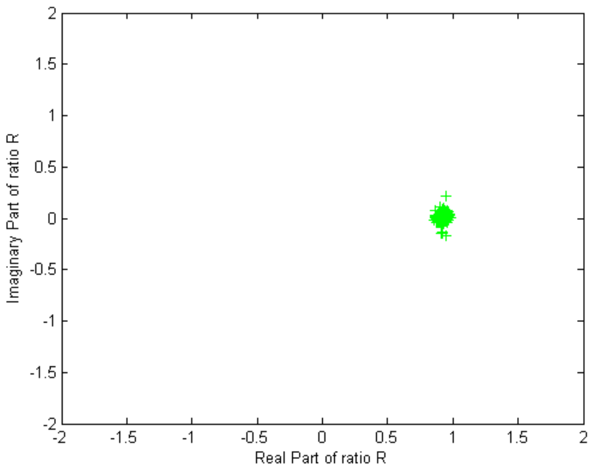

Figure 5.5: Ratios R for different downlink channel estimates

- $h_{d,i}$ are the downlink channels $i = 1, 2$,
- $h_{u,i}$ are the uplink channels $i = 1, 2$,
- $e^{j\phi}$ is the common phase difference between the uplink and the downlink channels.

Let $r_{t_1,t_2,i}$ be the ratio:

$$r_{t_1,t_2,i} = \frac{h_{u_{t_1},i}}{h_{d_{t_2},i}}, \quad i = 1, 2.$$

Considering the time slots t_i and t_j, we define the ratio $R_{i,j}$ as follows:

$$R_{i,j} = \frac{r_{t_1,t_2,1}}{r_{t_1,t_2,2}} = \frac{\dfrac{h_{u_{t_i},1}}{h_{d_{t_j},1}}}{\dfrac{h_{u_{t_i},2}}{h_{d_{t_j},2}}}.$$

If expression (5.3) holds true, then $R = 1$ (reciprocal channels), otherwise $R \neq 1$ (non-reciprocal channels).

Experiments

In this section, we use the uplink and downlink channel estimates at different time slots, and check if there is a common phase different between them. We follow the procedure described in the previous section, and calculate the ratios $R_{i,j}$ for various t_i, t_j. In Figure 5.6, we plot the ratios $R_{i,j}$ for $i = 1, 3, 5, ..., 99$ and $j = 2, 4, 6, .., 100$ (or, more precisely $R_{1,2}, R_{3,4}, .., R_{99,100}$).

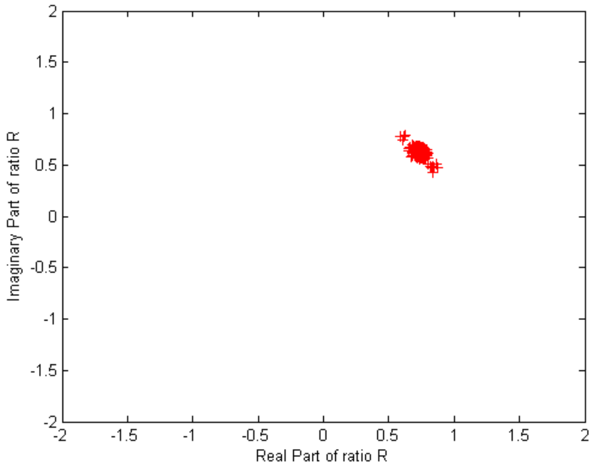

Figure 5.6: Ratios R for uplink and downlink channel estimations

We observe that the ratios $R_{i,j}$ are not located close to the point (1,0). This means that the 2x1 vector channels have not a common phase difference, i.e. expression (5.3) does not hold true.

Miscalibration

The fact that the reciproperty property does not hold true may be attributed to the non-symmetric characteristics of the RF electronic circuits.

Chapter 6

Interference Cancellation

In this chapter, we study the interference of wireless communication systems from "unwanted" sources. Let us assume two wireless communication systems, TX_1 - RX_1 and TX_2 - RX_2, that communicate simultaneously over the same frequency band, as depicted in Figure 6.1. Indeed, TX_1 interferes RX_2, and TX_2

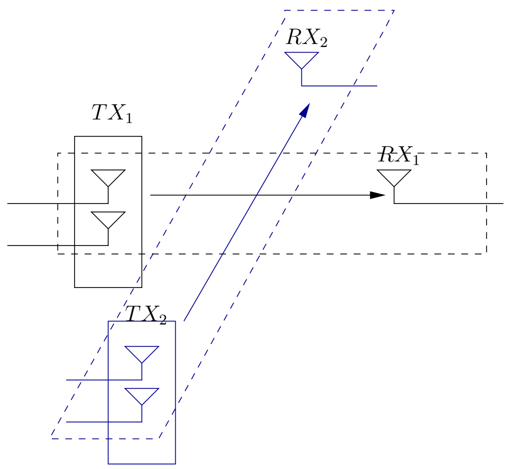

Figure 6.1: Simultaneous communication of two systems

interferes RX_1. In order to avoid this kind of interference, we suggest and study **beamforming**. We will consider these issues in detail and discuss our results.

6.1 The Problem

Let TX_1, and RX_1 be a communication pair which exchanges packets, and RX_2 a receiver which is interfered from TX_1 (assuming that they communicate in the same frequency band). Also, denote as s_1 and s_2 the signals of TX_1 which are transmitted from antenna 1 and antenna 2, respectively, Let the two channels between TX_1 and RX_2 be h_1 and h_2 as depicted in Figure 6.2. RX_2 receives the following signal (*interference*) from TX_1:

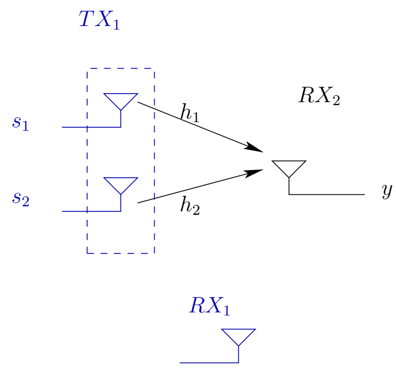

Figure 6.2: Interference from an unwanted source

$$\boxed{y = h_1 s_1 + h_2 s_2 + n}$$

where n is Additive White Gaussian Noise (AWGN).

6.2 Orthogonal Beamforming

Generally, beamforming is a signal processing technique used for directional signal transmission or reception.

Since our channels are not reciprocal, we cannot obtain the uplink channels, from the (estimated) downlink channels. Thus, we exploit the channel feedback in order to achieve beamforming and finally interference cancellation. Let $h_{1,n}$ and $h_{2,n}$ be the two channels between TX_1 and RX_2 at the n^{th} time slot, and let TX_1 send $s_1 = h_{2,n-1}s$ and $s_2 = -h_{1,n-1}s$, where s is the communication symbol

sequence.
We showed in the previous chapter, that there is a common phase difference $e^{j\phi}$, which satisfies the following expression:

$$\begin{bmatrix} h_{1,n} \\ h_{2,n} \end{bmatrix} = e^{j\phi} \begin{bmatrix} h_{1,n-1} \\ h_{2,n-1} \end{bmatrix}, \tag{6.1}$$

where $h_{i,n}$ are the uplink (or downlink) channels for $i = 1, 2$, at the time slot n. Hence, RX_2 receives:

$$\begin{aligned} y &= h_{1,n}s_1 + h_{2,n}s_2 + n \\ &= h_{1,n}h_{2,n-1}s + h_{2,n}(-h_{1,n-1}s) + n \\ &= (e^{j\phi}h_{1,n-1})h_{2,n-1}s + (e^{j\phi}h_{2,n-1})(-h_{1,n-1})s + n \\ &= e^{j\phi}(h_{1,n-1}h_{2,n-1} - h_{1,n-1}h_{2,n-1})s + n \\ &= 0 + n \\ &= n, \end{aligned}$$

where n is Additive White Gaussian Noise (AWGN). Thus, we showed how can TX_1 use orthogonal beamforming, in order to avoid the induced interference on RX_2.

6.3 Experiments

In this section, we discuss three different experiments regarding the interference cancellation.

6.3.1 A point to point system with an unintended receiver

We consider a system with three nodes; one pair of nodes, which consists of a two-antenna transmitter TX_1 and a single-antenna receiver RX_1, and one unintended receiver RX_2.
The experiment: At first, TX_1 and RX_2 communicate in order to estimate their channels and RX_2 sends its channel estimates ($h_{TX_1 \to RX_2}$) to TX_1. Then, TX_1 forms its beamformed signal, as we showed in the previous section, and sends N=100 beamformed packets for two different scenarios: (i) static channels (the

environment between USRPs remains constant), and (ii) non-static channels (the environment changes due to obstacle or TX, RX movements). In the static case, RX_1 receives most of the packets, whereas the unintended receiver RX_2 receives none. However, in the non-static case, RX_2, receives packets because the channel $h_{TX_1 \to RX_2}$ has changed. If we want to achieve again interference cancellation to the unintended receiver, we should re-estimate the *new* channels and re-create the new beamformed packets. Also, we have to mention that RX_1 receives most of the packets, as well as in the previous case. These results, are depicted in Figure 6.3.

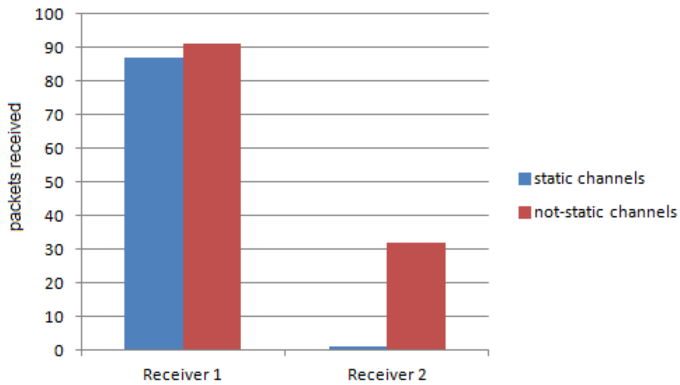

Figure 6.3: Number of packets received at the receivers for static and non-static channels.

6.3.2 Transmit and Orthogonal Beamforming

Let TX and RX be a two-antenna transmitter and a single-antenna receiver, respectively. In this experiment, the transmitted packet consists of two different beamformed sub-packets:

Transmit Beamforming

Regarding the first beamformed sub-packet (if $N_{symbols} = 100$, the first sub-packet consists of the $\frac{N_{symbols}}{2} = 50$ first symbols of the 100 overall symbols) we let TX_1 send $s_1 = h^*_{1,n-1}s$ and $s_2 = h^*_{2,n-1}s$, where s is the communication symbol sequence. We showed in the previous chapter, that there is a common phase difference $e^{j\phi}$, which satisfies the following expression:

$$\begin{bmatrix} h_{1,n} \\ h_{2,n} \end{bmatrix} = e^{j\phi} \begin{bmatrix} h_{1,n-1} \\ h_{2,n-1} \end{bmatrix}, \tag{6.2}$$

where $h_{i,n}$ are the uplink (or downlink) channels for $i = 1, 2$, at the time slot n. Thus, the first demodulated half part of the received packet is:

$$\begin{aligned} y &= h_{1,n}s_1 + h_{2,n}s_2 + n \\ &= h_{1,n}h^*_{1,n-1}s + h_{2,n}(h^*_{2,n-1}s) + n \\ &= (e^{j\phi}h_{1,n-1})h_{1,n-1}s + (e^{j\phi}h_{2,n-1})h^*_{2,n-1}s + n \\ &= e^{j\phi}(||h_1||^2 + ||h_2||^2)s + n, \end{aligned}$$

where n is Additive White Gaussian Noise (AWGN).

Orthogonal Beamforming

The second beamformed sub-packet (if $\frac{N_{symbols}}{2} = 50$, the second sub-packet consists of the $\frac{N_{symbols}}{2} = 50$ latter symbols of the 100 overall symbols) is the orthogonal beamformed sequence described in the previous section. Thus, the second half part of the received packet of RX is:

$$\begin{aligned} y &= h_{1,n}s_1 + h_{2,n}s_2 + n \\ &= h_{1,n}h_{2,n-1}s + h_{2,n}(-h_{1,n-1}s) + n \\ &= (e^{j\phi}h_{1,n-1})h_{2,n-1}s + (e^{j\phi}h_{2,n-1})(-h_{1,n-1})s + n \\ &= e^{j\phi}(h_{1,n-1}h_{2,n-1} - h_{1,n-1}h_{2,n-1})s + n \\ &= 0 + n \\ &= n, \end{aligned}$$

where n is Additive White Gaussian Noise (AWGN).
In this experiment, we observe that indeed, the second (orthogonal beamformed) half of the packet is utterly attenuated compared with the first (transmit beamformed), as depicted in Figure 6.4.

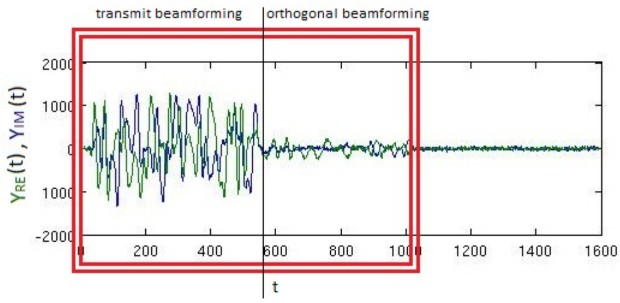

Figure 6.4: Transmit and Orthogonal Beamforming

6.3.3 Two 2x1 co-existing and interfering systems

We consider two 2x1 (two-antenna transmitter and a single-antenna receiver) systems, which communicate simultaneously over the same frequency band (Figure 6.1). Thus, TX_1 wants to communicate with RX_1 and TX_2 with RX_2. It is obvious that RX_1 and RX_2 are interfered from TX_2 and TX_1, respectively.

The experiment: First of all, TX_1 and RX_2 communicate in order to estimate their channels and RX_2 sends its channel estimations ($h_{TX_1 \to RX_2}$) to TX_1. Then, TX_1 forms its beamformed signal. Then, TX_2 and RX_1 communicate in order to estimate their channels and RX_1 sends its channel estimations ($h_{TX_2 \to RX_1}$) to TX_2. Afterwards, TX_2 forms its beamformed signal as well. Finally, we let both transmitters (TX_1 and TX_2) send N=100 beamformed packets for two different scenarios: (i) static channels (the environment between USRPs remains constant), and (ii) non-static channels (the environment changes due to obstacle or TX, RX movements)

Regarding the static case : On the one hand, RX_1 receives most of the packets transmitted from TX_1, and RX_2 receives most of the packets transmitted from TX_2. On the other hand, RX_1 does not receive packets transmitted from TX_2, and RX_2 does not receive packets transmitted from TX_1. However, *regarding the non-static case*: both receivers, receive packets from both transmitters, because the channels changes. If we want to achieve again interference cancellation to the unintended receivers, we should re-estimate the *new* channels and re-create the new beamformed packets. These results are depicted in Figures 6.4 (regarding transmitted packets from TX_1) and 6.5 (regarding transmitted packets from TX_2).

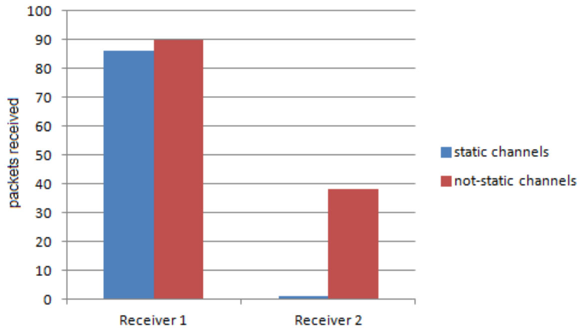

Figure 6.5: Number of packets sent from the first transmitter TX_1, and received at the receivers for static and non-static channels.

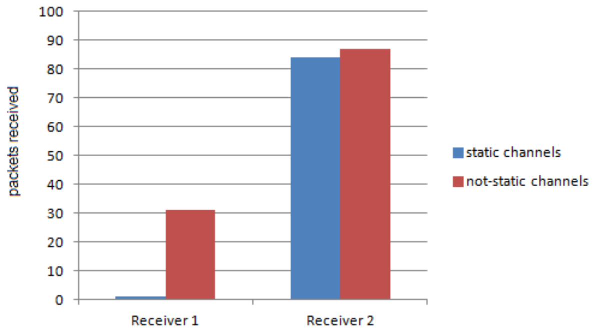

Figure 6.6: Number of packets sent from the second transmitter TX_2, and received at the receivers for static and non-static channels.

6.4 Generalization of "Orthogonal" Beamforming for N-1 receivers

We showed how can a single antenna receiver avoid the interference from a two-antenna transmitter. We can generalize this interference cancellation scheme, for an N-1 antenna receiver (or equivalently at N-1 distinct receivers), if we use an N-antenna transmitter. This generalization is shown in the Figure 6.6.

Let $\underline{X} = [\ x_1\ x_2\ ...\ \ x_n]^T$, or equivantly $\underline{X} = \underline{W}s$ be the transmitted vector, where s are the symbols, W the beamforming vector and $\underline{Y} = [\ y_1\ y_2\ ...\ y_{n-1}]^T$ the received vector at the receiver. Also, let $\underline{H}$ be the channel matrix:

$$\underline{H} = \begin{pmatrix} h_{1,1} & h_{1,2} & \cdots & h_{1,n} \\ h_{2,1} & h_{2,2} & \cdots & h_{2,n} \\ \vdots & \vdots & \ddots & \vdots \\ h_{n-1,1} & h_{n-1,2} & \cdots & h_{n-1,n} \end{pmatrix}.$$

We assume that H is full-rank. That is

$$\text{rank}(H) = n - 1.$$

We can cancel the interference at the N-1 receivers simultaneously i.e.

$$Y_m = 0, \quad m = 0, 1, .., n - 1,$$

or equivalently $\underline{Y} = \underline{HW}s = \underline{0} \Leftrightarrow \underline{HW} = \underline{0}$.
A solution to this problem is the eigenvector $H^H H$ associated with the zero eigenvalue.

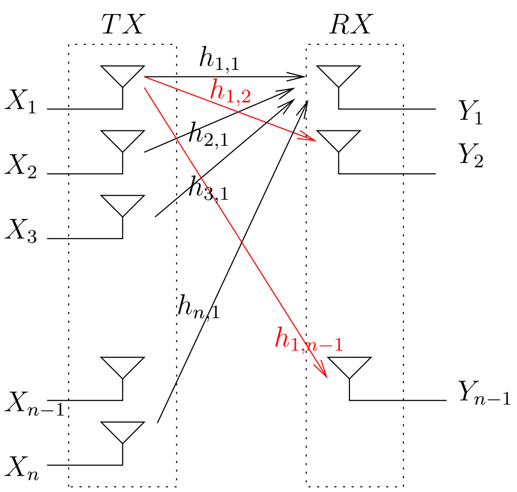

Figure 6.7: Interference Cancellation at N-1 receivers

Chapter 7

Conclusion

7.1 Conclusion

We showed how we can exploit Channel Feedback and the common phase channel difference in order to cancel the induced interference of two parallel wireless point to point systems. This work, can be generalized using a N-antenna transmitter and N-1 receivers.

7.2 Future Work

As future work, it can be suggested a calibration of the electronics of USRPs, in order to achieve reciprocal channels. Also, an implementation of a 3-antenna transmitter, which can send beamormed sequences at 2 different receivers.

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